AMPLIFIER LINEARIZATION BY PRE-DISTORTING A DECOMPOSED INPUT SIGNAL

Field of the Invention

The present invention relates to signal processing, and, in particular, to techniques for linearizing amplifiers based on input signal pre-distortion.

Background of the Invention

There are primarily two types of pre-distortion techniques used to linearize power amplifiers: digital and analog. Digital techniques typically use envelope detectors, analog-to-digital converters (ADCs), look-up-tables, digital-to-analog converters (DACs), and vector modulators to create the pre-distorted waveforms to be applied to the power amplifier. Within reason, a multitude of pre-distorted waveforms can be created with this type of technique. However, such digital techniques include an inherent throughput delay associated with the ADC, digital signal processing, and DAC circuits. Most of this delay is related to the clock speed of operation, since each of these circuits is typically associated with many clock cycles of delay. To fundamentally reduce the delay, the clock speed must be increased. Additionally, digital techniques usually require expensive ADC and DAC components in the signal processing chain.

Existing analog predistortion circuits generally have fundamental limitations, since precise amplitude/phase relationships are not easily derived. Most analog circuits used are diode- or transistor-based circuits that do not have independent gain and phase relationships or controls. This means that a particular non-linear element is used for pre-distortion and is merely an approximation to the ultimate desired transfer curve. For example, a particular non-linear element, e.g., a transistor, is used as the pre-distortion device. The device may be driven hard enough to create the desired amplitude compression that matches the power amplifier's compression curve. However, as the pre-distorter is driven hard, the phase also compresses, but to a degree that may or may not match the power amplifier's phase-compression curve, resulting in a mismatch between the pre-distorter and the power amplifier, reducing the effectiveness of the pre-distorter. If the phase does match, it is only under very unique conditions that might not apply to another power amplifier's transfer curve, producing repeatability problems. As a result, most analog pre-distorters produce inadequate approximations to the amplifier's compressive characteristics.

Brief Description of the Drawings

Aspects, features, and advantages of the present invention will become more fully apparent from the following detailed description, the appended claims, and the accompanying drawings in which like reference numerals identify similar or identical elements.

Figs. 1 and 2 graphically represent typical amplitude and phase transfer curves, respectively, for an LDMOS-based class AB power amplifier;

Fig. 3 shows a block diagram illustrating amplitude pre-distortion of an input signal, according to one embodiment of the present invention;

Fig. 4 shows a block diagram illustrating phase pre-distortion of an input signal, according to one embodiment of the present invention;

Fig. 5 shows a schematic diagram of a non-linear circuit designed for second-quadrant operations, while Fig. 6 shows a schematic diagram of a non-linear circuit designed for fourth-quadrant operations;

Fig. 7 shows a schematic diagram of a common differentiator circuit implemented using an op-amp;

Fig. 8 shows a block diagram illustrating frequency-dependent amplitude pre-distortion of an input signal, according to one embodiment of the present invention; and

Fig. 9 shows a block diagram illustrating frequency-dependent amplitude pre-distortion of an input signal, according to another embodiment of the present invention.

Detailed Description

Figs. 1 and 2 graphically represent typical amplitude and phase transfer curves, respectively, for an LDMOS-based class AB power amplifier. The normalized amplitude transfer curve of Fig. 1 nominally consists of two regions: a gain-expansion region generally at low input signal levels (e.g., less than an index of 200 in Fig. 1) and a gain-compression region generally during the signal peaks or at high input signal levels (e.g., greater than an index of 200). The amplitude curve shown in Fig. 1 is a 3rd-order transfer function, although typical power amplifiers can have higher-order transfer functions, particularly noticeable at the signal peaks. The phase curve shown in Fig. 2, on the other hand, is a 2nd-order polynomial, by way of example. The present invention is not limited by the order of the transfer curves and can be applied to a wide range of relationships exhibiting the general shape shown in these figures.

Amplitude Pre-Distortion

Fig. 3 shows a block diagram illustrating amplitude pre-distortion of an input signal, according to one embodiment of the present invention. As shown in Fig. 3, an analog RF input signal 302 is delayed by delay element 304 and then amplitude-adjusted at amplitude modulator 306 (e.g., a voltage-controlled variable attenuator) to generate an amplitude-distorted RF output signal 308.

A portion of RF input signal 302 is coupled off by tap 310, optionally attenuated at attenuator 312, and applied to diode 314, which functions as an envelope detector that provides an instantaneous measure or indication of the input signal amplitude. After the signal is envelope detected, the voltage waveform 316 is shown as a two-tone signal from 0 volts to Vmax, where 0 volts represents the signal-envelope nulls and Vmax represents the signal-envelope peaks.

This signal is split (318) into two copies, one of which is applied to DC-level-shifting circuit 320, nominally created using op-amps, which shifts the applied signal such that the signal-envelope peaks are referenced to 0 volts and the signal-envelope nulls are referenced to Vmin, as shown in voltage waveform 322. This DC-level-shifted signal is input to a first non-linear transfer function circuit 324, which operates in quadrant II in a Cartesian coordinate system, as shown in graph 326. The signal-envelope peaks (aligned with 0 volts in waveform 322) will be largely unaffected by circuit 324, since the output voltage in this region is near 0 volts. Only the large negative voltages, representing the signal-envelope nulls (or low power) will be affected by circuit 324. In this case, a positive voltage is produced for these lower signal levels.

The other copy of the envelope-detected signal is applied (without a DC shift) to a second non-linear transfer function circuit 328, which operates in quadrant IV, as shown in graph 330. Circuit 328 is designed such that low input voltages, near 0 volts, nominally produce 0-volt output voltages, while the larger input voltages produce negative output voltages at high levels, as shaped by the transfer function. Circuit 328, therefore, operates mainly on the signal-envelope peaks and leaves the signal-envelope nulls largely unaffected.

The signals produced by circuits 324 and 328 are combined at summation node 336 to create a complete transfer function, represented by waveform 338. In particular, circuits 324 and 328 are designed such that waveform 338 mimics the gain expansion and compression characteristics, respectively, of the amplifier (not shown) to be linearized. This combined signal is inverted by an inverting circuit 340 to change the sign of the transfer curve as represented by waveform 342.

The resulting inverted signal is used to control voltage-controlled attenuator 306, which adjusts the amplitude of the delayed RF input signal based on the input signal's instantaneous envelope value, where delay element 304 delays the main signal path an amount approximately equal to the analog signal

processing delay of the control signal path. In the example shown in Fig. 3, lower control voltage values (e.g., less than zero) adjust the attenuator to yield higher attenuation in the main signal path, consistent with a signal decrease used to compensate for amplifier gain expansion. Conversely, higher control voltage levels (e.g., greater than zero) adjust the attenuator to yield lower attenuation in the main signal path, consistent with a signal boost used to compensate for amplifier gain compression.

Fig. 3 indicates control signals being applied to non-linear circuits **324** and **328**. These control signals are intended to represent the performance of adaptation on the circuits' transfer curves to compensate for the effects of time, temperature, aging, and the like.

Phase Pre-Distortion

Fig. 4 shows a block diagram illustrating phase pre-distortion of an input signal, according to one embodiment of the present invention. The processing of Fig. 4 is analogous to that of Fig. 3, except that the analog circuits are designed to generate a combined transfer function that models the phase characteristics of the amplifier to be linearized rather than the amplifier's gain characteristics. In particular, as shown in Fig. 4, an analog RF input signal **402** is delayed by delay element **404** and then phase-adjusted at phase modulator **406** (e.g., a voltage-controlled phase shifter) to generate a phase-distorted RF output signal **408**.

A portion of RF input signal **402** is coupled off by tap **410**, optionally attenuated at attenuator **412**, and applied to diode **414**, which functions as an envelope detector. After the signal is envelope detected, the voltage waveform **416** is shown as a two-tone signal from 0 volts to Vmax, where 0 volts represents the signal-envelope nulls and Vmax represents the signal-envelope peaks.

This signal is split (**418**) into two copies, one of which is applied to DC-level-shifting circuit **420**, which shifts the applied signal such that the signal-envelope peaks are referenced to 0 volts and the signal-envelope nulls are referenced to Vmin, as shown in voltage waveform **422**. This DC-level-shifted signal is input to a first non-linear transfer function circuit **424**, which operates in quadrant II, as shown in graph **426**. The signal-envelope peaks (aligned with 0 volts in waveform **422**) will be largely unaffected by circuit **424**, since the output voltage in this region is near 0 volts. Only the large negative voltages, representing the signal-envelope nulls (or low power) will be affected by circuit **424**. In this case, a positive voltage is produced for these lower signal levels.

The other copy of the envelope-detected signal is applied (without a DC shift) to a second non-linear transfer function circuit **428**, which operates in quadrant IV, as shown in graph **430**. Circuit **428** is designed such that low input voltages, near 0 volts, nominally produce 0-volt output voltages, while the larger input voltages produce negative output voltages at high levels, as shaped by the transfer function.

Circuit **428**, therefore, operates mainly on the signal-envelope peaks and leaves the signal-envelope nulls largely unaffected. The output from circuit **428** is inverted by inverting circuit **432** to generate waveform **434**.

The signals produced by circuit **424** and inverter **432** are combined at summation node **436** to create a complete transfer function, represented by waveform **442**, where circuits **424** and **428** are designed such that waveform **442** mimics the phase characteristics of the amplifier (not shown) to be linearized. This combined signal is used to control voltage-controlled phase shifter **406**, which adjusts the phase of the delayed RF input signal based on the input signal's instantaneous envelope value, where delay element **404** delays the main signal path an amount approximately equal to the analog signal processing delay of the control signal path. In the example shown in Fig. 4, increasing control voltage magnitude (i.e., on either side of zero) corresponds to increased insertion phase in the main signal path. A phase pre-distortion of opposite signal could be handled using an inverting op-amp circuit in the control signal path.

As in Fig. 3, the control signals shown in Fig. 4 represent the performance of adaptation on the circuits' transfer curves to compensate for the effects of time, temperature, aging, and the like.

Combined Amplitude and Phase Pre-Distortion

Depending on the implementation, the amplitude and phase pre-distortion circuits of Figs. C and D can be combined either in series or in parallel to form a single complex amplitude-and-phase pre-distortion circuit that can be (i) relatively low in throughput delay, (ii) implemented using relatively low-cost components, and (iii) amenable to adaptation, which may be critical in certain high-performance applications.

When implemented in series (i.e., either amplitude before phase or vice versa), the RF output signal from the first pre-distortion circuits forms the RF input signal for the second pre-distortion circuit, where the RF output signal from the second pre-distortion circuit represents the amplitude-and-phase pre-distorted signal that may be applied to a corresponding (high power) amplifier.

When implemented in parallel, the circuits can share the same tap, delay element, and envelope detector circuitry, where the envelope-detected signal drives four non-linear transfer curve circuit paths, analogous to those of circuits **324** and **328** of Fig. 3 and circuits **424** and **428** of Fig. 4. The results are combined in appropriate pairs and used to drive a series combination of an attenuator (analogous to attenuator **306** of Fig. 3) and a phase adjuster (analogous to phase adjuster **406** of Fig. 4) to yield a complex, pre-distorted output signal.

Depending on the implementation, the circuit paths corresponding to non-linear circuits **324** and **424** can also share the same DC-level-shifting circuit, whose output is then split for application to the

different amplitude and phase transfer function circuits. Moreover, depending on the actual shapes of the amplitude and phase transfer functions, it may also be possible to share one or more non-linear circuits. For example, the transfer function of first non-linear circuit **324** of Fig. 3 and the transfer function of first non-linear circuit **424** of Fig. 4 appear to have similar shapes. If the differences are sufficiently small for a particular application, a single non-linear circuit could be used for both the amplitude and phase pre-distortion processing, with two different copies of the resulting waveform being applied to two different summation nodes, one for amplitude pre-distortion and the other for phase pre-distortion.

Each of non-linear analog circuits **324**, **328**, **424**, and **428** may be implemented using a piece-wise linear approximation of the transfer curve or a non-linear functional element yielding the desired curve (e.g., a diode or several diodes in series or a transistor with a desirable turn-on characteristic).

Fig. 5 shows a schematic diagram of a non-linear circuit designed for second-quadrant operations that could be used for non-linear circuits **324** and **424**, while Fig. 6 shows a schematic diagram of a non-linear circuit designed for fourth-quadrant operations that could be used for non-linear circuits **328** and **428**.

Frequency-Dependent Embodiments

The embodiments of the present invention shown in Figs. 3 and 4 correspond to frequency-independent pre-distorters (referred to herein as "layer 1" pre-distortion). The concept of the present invention can be extended such that differentiation is applied to the pre-distorted signal to produce a pre-distorted signal that is dependent on frequency (referred to herein as "layer 2" pre-distortion). The differentiated, layer 2 pre-distorted signal is combined with the original, layer 1 pre-distorted signal to produce a pre-distorted signal having both frequency-independent and frequency-dependent characteristics. The "memory effects" prevalent in most class AB amplifiers can be modeled or approximated by a differentiator circuit, which has frequency-dependent amplitude and phase characteristics.

Fig. 7 shows a schematic diagram of a common differentiator circuit **700** implemented using an op-amp **702**. In circuit **700**, the output voltage Vout is related to the input voltage Vs according to Vout = -RC d/dt{Vs}. This is equivalent to stating that the gain Gs is $-j\omega RC$, where ω is the angular frequency equal to $2\pi f$. These relationships indicate the frequency-dependent and signal-inversion characteristics of the circuit as well as its gain of zero at zero frequency.

Fig. 8 shows a block diagram illustrating frequency-dependent amplitude pre-distortion of an input signal, according to one embodiment of the present invention. The pre-distortion processing of Fig. 8 is similar to that of Fig. 3, except that, in addition to inverter **840** (which is analogous to inverter **340** of Fig. 3), Fig. 8 has an inverting differentiator **844** and a summation node **846**. As shown in Fig. 8, in

addition to being applied to inverter **840**, the combined waveform **838** is applied to inverting

differentiator **844**, which may be based on differentiator circuit **700** of Fig. 7. The resulting inverted

waveform from inverter **840** and the resulting inverted and differentiated waveform from differentiator

844 are combined at summation node **846** to form a combined waveform **842**, which is applied to an

5   attenuator analogous to attenuator **306** of Fig. 3 to generate a pre-distorted signal for subsequent

amplification. In this embodiment, the two non-linear circuits **824** and **828** are each used for both the

frequency-independent and the frequency-dependent components, thereby conserving circuitry.

        Fig. 9 shows a block diagram illustrating frequency-dependent amplitude pre-distortion of an

input signal, according to another embodiment of the present invention. The pre-distortion processing of

10   Fig. 9 is similar to that of Fig. 8, except that, in Fig. 9, separate circuitry is provided for generating the

frequency-independent and the frequency-dependent components, thereby increasing the flexibility of

this pre-distortion processing. In particular, Fig. 9 shows two DC shifters **920i** and **920d**, two second-

quadrant non-linear circuits **924i** and **924d**, two fourth-quadrant non-linear circuits **928i** and **928d**, and

two summation nodes **936i** and **936d**, where the suffix "i" identifies a frequency-independent component

15   and the suffix "d" identifies a frequency-dependent component. As such, inverter **940** and inverting

differentiator **944** receive two potentially different, combined waveforms, respectively, the former

corresponding to frequency-independent characteristics and the latter corresponding to frequency-

dependent characteristics.

        By extension and analogous to Fig. 8, frequency-dependent components can also be added to the

20   frequency-independent phase pre-distortion processing of Fig. 4. In particular, and referring to Fig. 4,

after summation node **436**, a splitter can be added followed (in one of the splitter's downstream paths) by

either a non-inverting differentiator or a combination of an inverting differentiator and an inverter, to

generate a non-inverted, differentiated (i.e., frequency-dependent) pre-distortion signal that is then added

to the original, frequency-independent waveform **442** to form a combined waveform having both

25   frequency-independent and frequency-dependent characteristics. Alternatively, a second set of

frequency-dependent circuitry could be added analogous to Fig. 9.

Alternative Embodiments

        The present invention has been described in the context of non-linear circuits that operate in

30   quadrants II and IV (e.g., circuits **324** and **328** of Fig. 3 and circuits **424** and **428** of Fig. 4). This

corresponds to preferred implementations in which inverting op-amp circuits, which always operate in

quadrant II or IV, are used to keep the gain less than one. In theory, however, the present invention could

be implemented using one or more non-linear circuits that operate in either quadrant I or III. Depending

on the implementation details, such embodiments might or might not need inverters analogous to inverters **340** of Fig. 3 and **432** of Fig. 4.

Although the present invention has been described in the context of pre-distortion of an LDMOS-based class AB power amplifier, the invention is not so limited. In general, the present invention can be

5   implemented to pre-distort signals input to any type of amplifier exhibiting suitable compression characteristics.

Although the present invention has been described in the context of amplitude and phase pre-distortion processing in which an input signal is decomposed and used to generate two different halves of each transfer function, the invention is not so limited. In general, the present invention can be

10   implemented to model two or more different portions of the amplitude and/or phase transfer functions of an amplifier to be linearized.

The present invention may be implemented in the context of wireless signals transmitted from a base station to one or more mobile units of a wireless communication network. In theory, embodiments of the present invention could be implemented for wireless signals transmitted from a mobile unit to one

15   or more base stations. The present invention can also be implemented in the context of other wireless and even wired communication networks to reduce spurious emissions.

Embodiments of the present invention may be implemented as circuit-based processes, including possible implementation on a single integrated circuit (such as an ASIC or an FPGA), a multi-chip module, a single card, or a multi-card circuit pack.

20   It will be further understood that various changes in the details, materials, and arrangements of the parts which have been described and illustrated in order to explain the nature of this invention may be made by those skilled in the art without departing from the scope of the invention as expressed in the following claims.